Integrated Millimeter-Wave Corner-Cube Antennas

Steven S. Gearhart, *Student Member, IEEE,* Curtis C. Ling, *Student Member, IEEE,* and Gabriel M. Rebeiz, *Member, IEEE*

Abstract—**An integrated corner-reflector antenna has been designed, fabricated and measured at millimeter-wave frequencies. The structure consists of a traveling-wave antenna integrated on a 1.2-μm dielectric membrane and suspended in a longitudinal cavity etched in a silicon wafer. A new traveling wave antenna design, the modified-bend antenna, with an antenna length of $L = 1.2\lambda$ and spacing $d = 0.96\lambda$ from the apex, results in a wide-band input impedance centered at 140 Ω and low cross-polarization levels. Measurements at 180–270 GHz show a well defined pattern with low sidelobe levels and a main-beam efficiency of 93 and 83% at 180 and 222 GHz, respectively. The monolithic approach allows the integration of a matching network and a Schottky-diode or SIS detector at the base of the antenna to yield a low-noise monolithic millimeter-wave receiver.**

I. Introduction

THE standard corner-cube antenna, which consists of a traveling-wave antenna backed by a 90° corner reflector [1]–[7], has traditionally been the workhorse of submillimeter-wave receivers. The corner-cube is an open structure mixer mount, and can be made to operate at much higher frequencies than the waveguide mixer. The standard design is a 4λ-long traveling-wave antenna placed 1.2λ from the apex of the reflector. The antenna also acts as a whisker contact to a Schottky-diode mounted at its base. It is a high gain antenna and couples well to $f/1.5$-1.8 imaging systems [7]. Although the corner-cube has been considered as the "best" available antenna at these frequencies, it has a number of serious problems. First, the corner-cube antenna cannot be readily modified to include an RF matching network. The antenna impedance is around 160 Ω, and therefore couples poorly to the Schottky-diode which requires an optimum RF-impedance of about $60 + j70$ Ω [8]. The antenna has high sidelobes and a relatively high cross-polarization component, and typically about 50–60% of the radiated power is coupled into a fundamental Gaussian beam [1], [6], [7]. Although no accurate coupling-efficiency measurements have been performed on the corner-cube receiver, it is currently believed that a value less than 25% (this includes main-beam efficiency and impedance matching to the diode) is achieved at submillimeter-wave frequencies. The poor coupling efficiency contributes directly to the high-noise temperature (17 000–25 000°K-SSB-cooled) of corner-cube receivers available at 2000–2500 GHz [9], [10].

To solve the problems associated with the machined corner-cube antenna, an integrated version has been developed at the University of Michigan [11], [12]. The structure consists of a traveling-wave antenna suspended on a 1.2-μm dielectric membrane in a longitudinal pyramidal cavity (Fig. 1). The membrane electrical thickness is 0.02λ at 3 THz, and the traveling-wave antenna effectively radiates in free space at all frequencies of interest. The cavity is etched in silicon wafers [13], and the flare angle is fixed by the orientation of the crystal planes at 70.6°. It is possible to achieve this angle, and many other interesting angles such as 60° and 90°, using anisotropic etching of GaAs [14]. This is important because one could integrate these antennas with planar GaAs Schottky-diode detectors. The monolithic approach allows the integration of a complete receiver with the RF and IF matching networks and is expected to cost much less than the standard machined version, particularly in arrays with a large number of elements.

II. Design of the Integrated Corner-Cube Antenna

The theoretical pattern of a corner-reflector element is calculated by first assuming the reflecting ground planes to be infinite in extent. The total pattern is the product of the element pattern and an array pattern. The element pattern is the pattern of a long-wire antenna. The antenna is assumed to have a constant traveling-wave current from the feed out to the tip of the antenna and zero current after the sharp discontinuity in the wire [15]. For the array pattern, the corner reflector cannot be modeled by the method of images since the subtended angle is not given by $180°/n$, where n is an integer. However, for the initial design, a 60° subtended angle is close enough to 70.6°, and the array pattern is calculated for a 60° corner reflector. Extensive microwave modeling was also performed on a corner-reflector with a 70.6° subtended angle. The radiation patterns of traveling-wave antennas with lengths between 1.0–1.5λ placed 0.85–1.05λ from the apex resulted in a well defined mainlobe and low sidelobe levels. A design with an $(L, d) = (1.2\lambda, 0.96\lambda)$ resulted in nearly equal E- and quasi-H 10-dB-beamwidth of 44° and a sidelobe level of −16 dB. The measured E-plane and quasi-H plane patterns agree fairly well with the patterns predicted using a 60° corner-reflector

Manuscript received September 11, 1990; revised January 28, 1991. This work was supported by the NASA/Center for Space Terahertz Technology, University of Michigan, Ann Arbor.

The authors are with the NASA/Center for Space Terahertz Technology, Electrical Engineering and Computer Science Department, University of Michigan, Ann Arbor, MI 48109-2122.

IEEE Log Number 9100569.

0018-926X/91/0700-1000$01.00

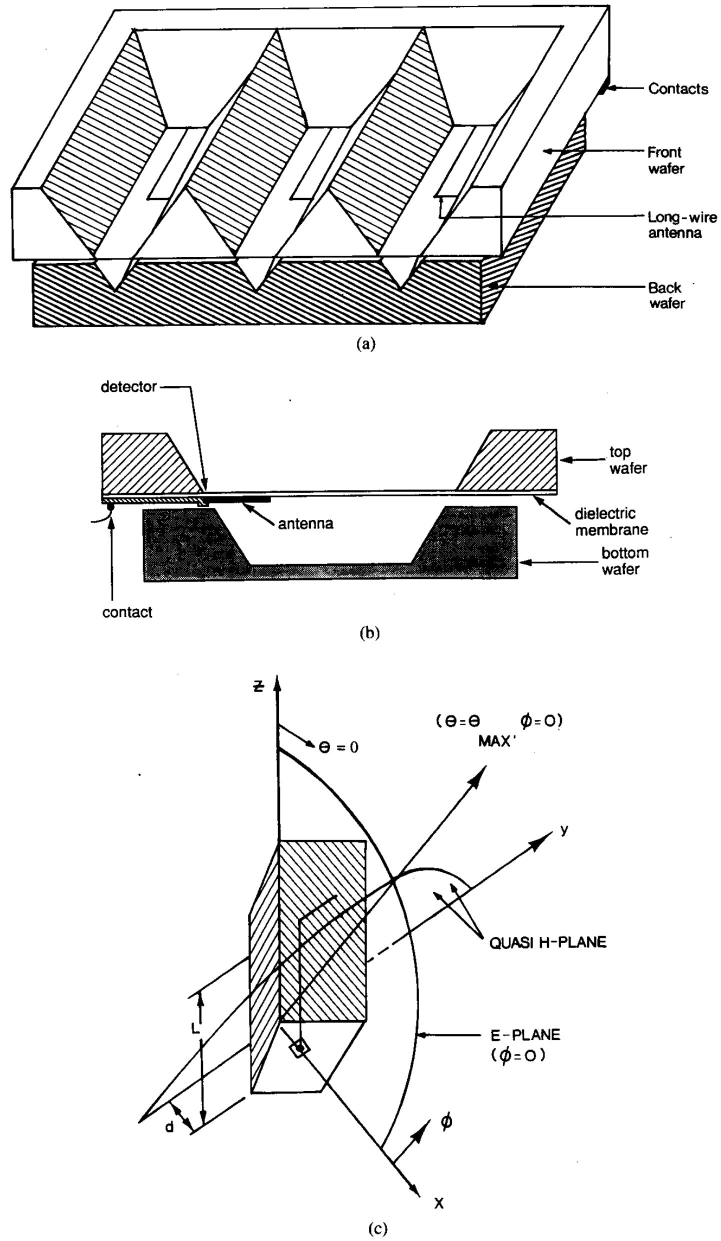

Fig. 1. A monolithic corner-reflector imaging array. (a) Perspective view. (b) Side view. (c) Coordinate system used.

(Fig. 2). The measured H-plane pattern is wider than the calculated one due to the larger corner-reflector angle, and the E-plane pattern is shifted by 5°. It is therefore clear that the theoretical model gives a good initial design, and one that can be used for first-order calculations.

The traveling-wave antenna with a simple bend shows a large cross-polarization component in the E-plane of -9.5 dB in microwave model measurements. Since the RF currents do not attenuate quickly after the sharp discontinuity, the RF currents in the bend contribute significantly to the cross-polarization component. One way to solve this problem is to reduce the width of the bend portion of the antenna. The current in the narrow bend does not contribute significantly to the cross-polarization component in the E-plane. This antenna is called the modified-bend antenna (Fig. 3). The modified-bend shows essentially the same E- and quasi-H plane patterns, but with a measured peak cross-polarization component in the E-plane of -16 dB. The cross-polarization component in the quasi-H plane is always present due to the off-broadside radiation of the corner reflector antenna and the

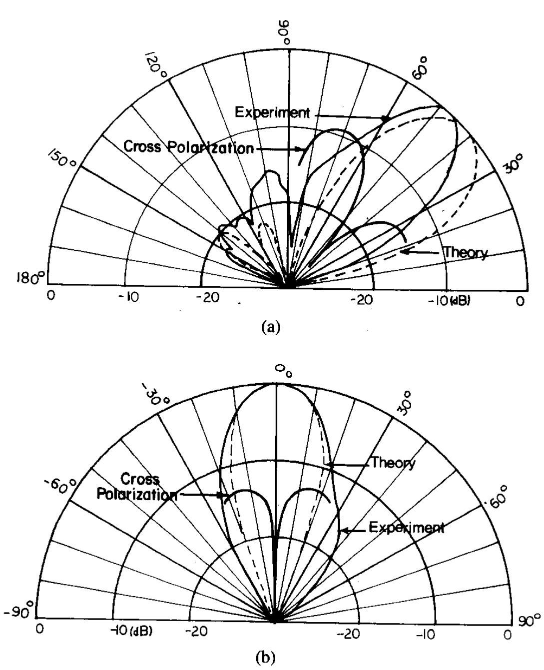

Fig. 2. Measured 3.3-GHz patterns and predicted E-plane (a) and quasi-H plane (b) patterns for a standard traveling-wave antenna and $(L, d) = (1.2\lambda, 0.96\lambda)$.

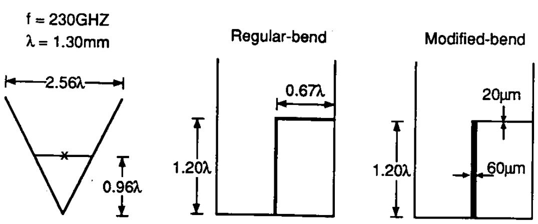

Fig. 3. Standard traveling-wave and modified-band antennas. The design is for a center frequency of 230 GHz.

measured peak values for the regular and modified-bend antennas are -13 and -15 dB, respectively.

The modified-bend antenna with an $(L, d) = (1.2\lambda, 0.96\lambda)$ was chosen for millimeter-wave integration. The measured input impedance at microwave frequencies is broad band with a radiation resistance centered at 140 Ω a reactance lower than 60 Ω (Fig. 4). The input impedance increased by 20 Ω for a narrower antenna (40 μm at 230 GHz). The impedance of the modified bend antenna is essentially identical to that of the standard traveling-wave antenna, and this shows that the narrow bend portion does not disturb significantly the traveling-wave current distribution.

Finally, a bandwidth study was done on the radiation patterns of the modified-bend antenna (Fig. 5). The antenna length and spacing (L, d) of $(1.2\lambda, 0.96\lambda)$ at 2.6 GHz (scale-model for the 230-GHz antenna) charges considerably for a 2.0 GHz to 3.0 GHz frequency span. A shift in the E-plane pattern peak of 20° occurred between 2.0 GHz and 3.0 GHz, and this is expected from a traveling-wave antenna.

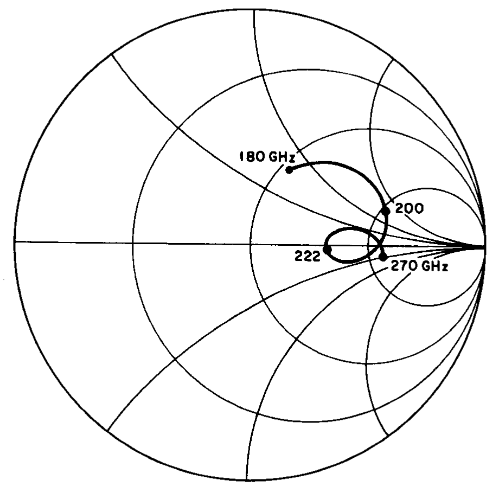

Fig. 4. The measured input impedance of the modified traveling-wave antenna. Measurements were done from 2–3 GHz, and the corresponding millimeter-wave frequencies shown.

It is clear that this design, in conjunction with the measured input impedance, results in a large bandwidth antenna. However, the axis of the corner-reflector should be aligned with the optical system to compensate for the shift in the E-plane pattern.

III. Fabrication

The corner-reflector antenna is a stacked silicon wafer structure. The thickness of the front wafer determines the position of the traveling-wave antenna inside the pyramidal cavity. It also determines the extension of the ground planes from the radiating element. The back wafer completes the pyramidal structure and acts as a reflecting cavity.

The pyramidal cavity is made by anisotropic etching of silicon wafers with a ⟨100⟩ crystallographic orientation [13]. This etching process naturally forms pyramidal holes bounded by the ⟨111⟩ crystal planes. The membrane layer is fabricated by depositing a three-layer SiO_2/Si_3N_4 structure which is in tension, and etching the underlying silicon until the transparent membrane appears. After etching, cavity walls are coated with gold using an angle evaporation technique, and the antennas, detectors and low-frequency lines are deposited on the back-side of the front wafer using standard lithography. Finally, the wafer stack is assembled by aligning the wafers and gluing them together. The alignment fixture is accurate to within ±5 μm and the glue gap is very small (estimated at 2–3 μm). A 119-μm integrated corner-cube antenna has been fabricated using this method. The patterns show equal E- and quasi-H plane patterns and agree well with 222-GHz patters [16].

The modified-bend antenna is 2000 Å thick and is one skin-depth thick at 230 GHz. The antenna is 1.56-mm long (1.2λ at 230 GHz), and is 60-μm wide with a 20-μm bend portion. The antenna is suspended on a 1.2-μm-thick membrane in a pyramidal cavity with an opening 3.3-mm-wide (2.55λ at 230 GHz) and 1-cm-long (7.7λ at 230 GHz). A

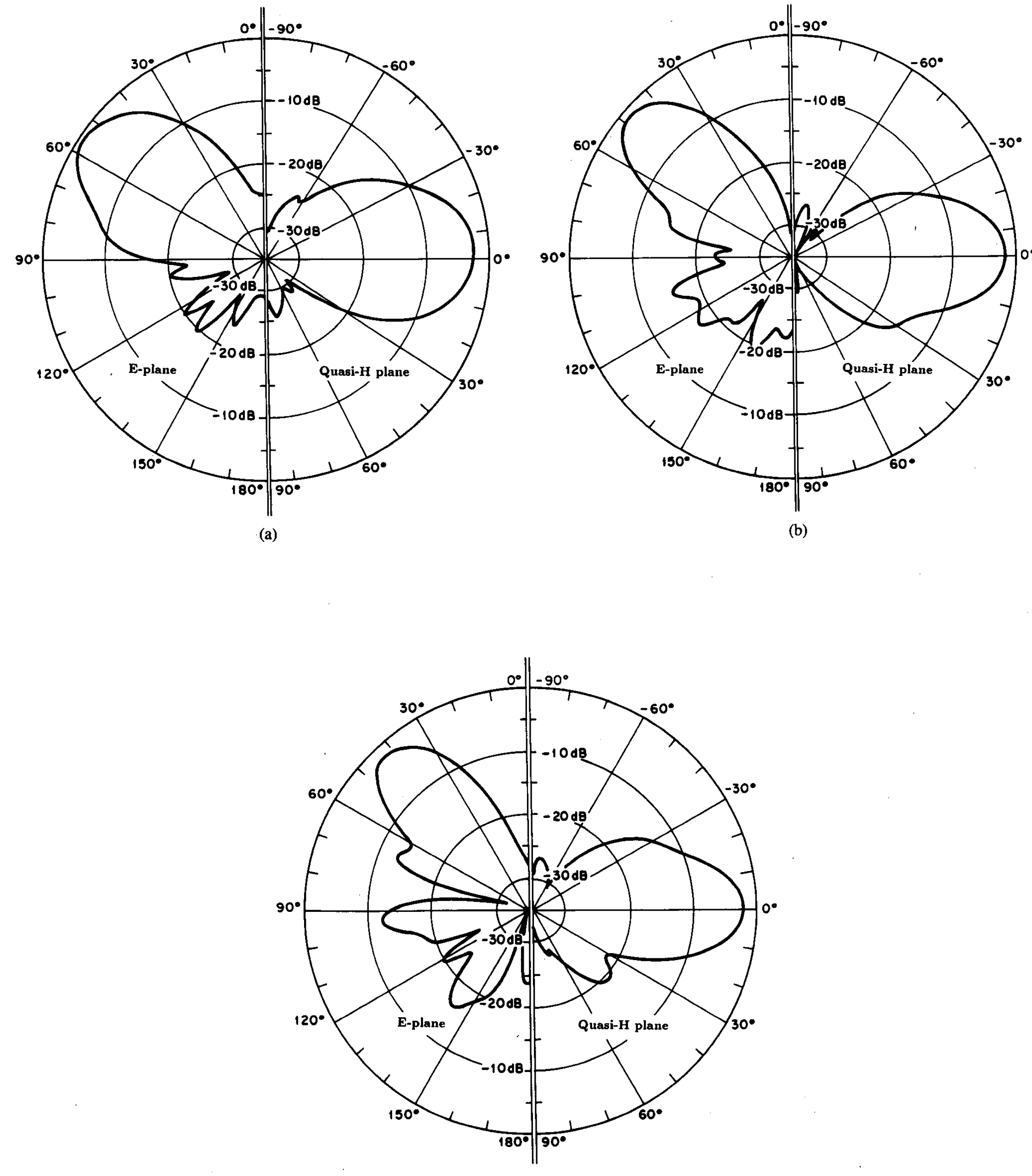

Fig. 5. *E*-plane and quasi-*H* plane pattern measurements at (a) 2.0 GHz, (b) 2.5 GHz; and (c) 2.8 GHz. Measurements (not shown) done at 2.2, 2.6, and 3.0 GHz show a similar trend and well behaved patterns.

4-μm-square microbolometer [17] was integrated at the bottom tip of the traveling-wave antenna. This is the same position that one would integrate a matching network and a Schottky-diode or SIS detector in a receiver application. The bolometer impedance was 80 Ω with a responsivity of 6 V/W at 0.1-V bias.

IV. Millimeter-Wave Measurements

The patterns were measured at 180, 222, 240, and 270 GHz using millimeter-wave doublers and triplers fed by appropriate Gunn sources. The sources were AM modulated at 300 Hz, and the output from the bolometers was feed to a PAR-124A lock-in amplifier. Pattern measurements done on

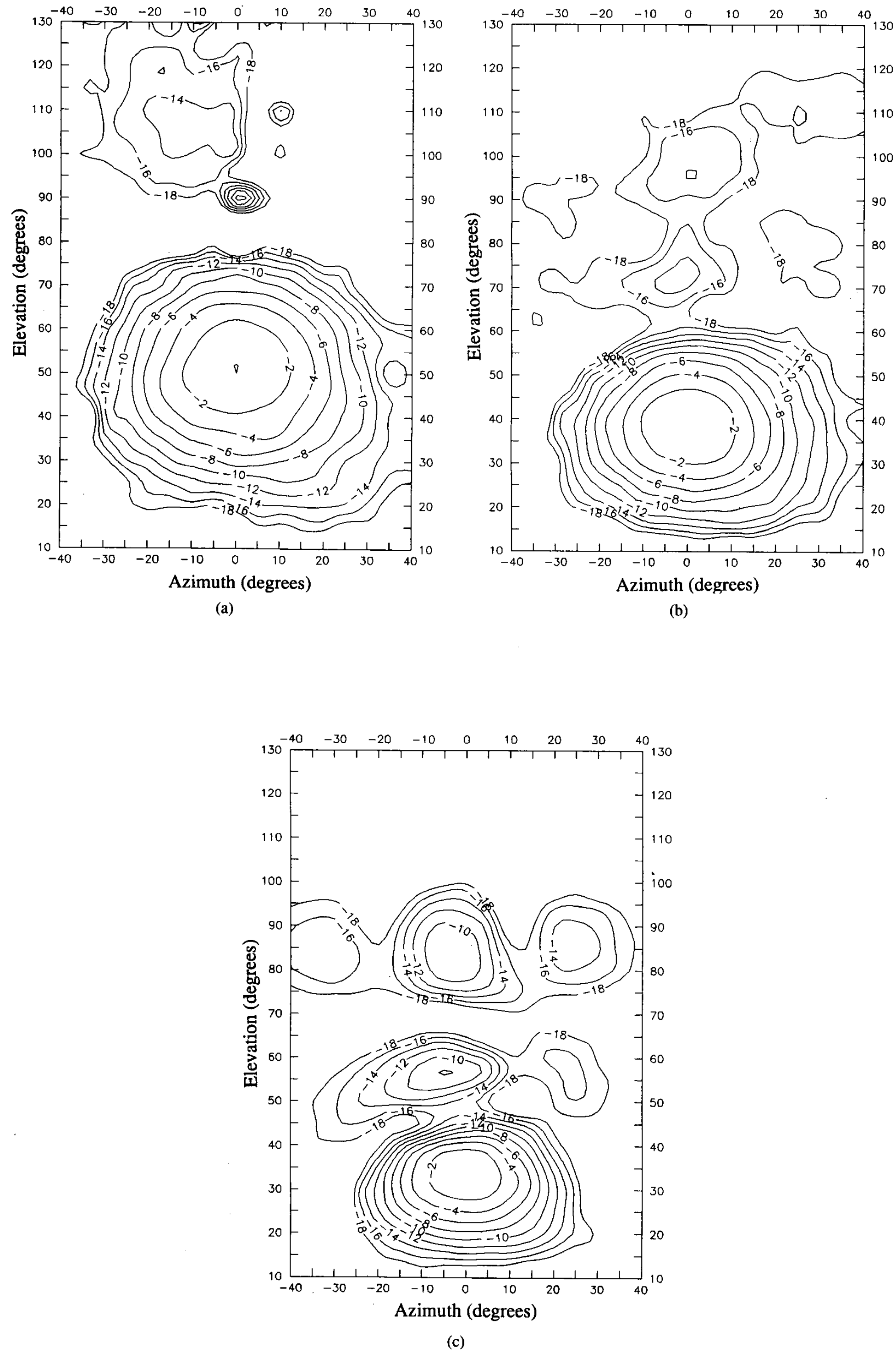

Fig. 6. Measured two-dimensional patterns on a 230 GHz design with an $(L, d) = (1.2\lambda, 0.96\lambda)$. Pattern measurements were done at (a) 180 GHz; (b) 222 GHz; and (c) 270 GHz. (See Table I).

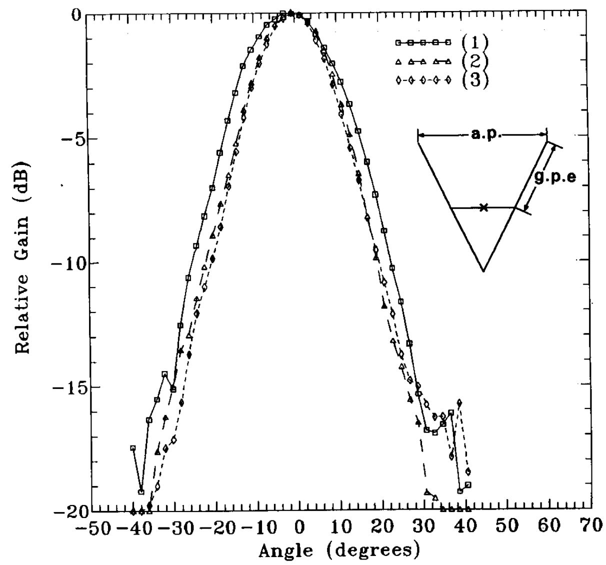

Fig. 7. Measured quasi-H plane patterns at 222 GHz for three identical antennas with ground-plane extensions (GPE) of 0.94λ-(1), 1.6λ(2), and 2.3λ(3) from the antenna, yielding a array period (AP) of 2.5λ, 3.2λ, and 4.0λ, respectively.

TABLE I

Frequency	180 GHz	222 GHz		270 GHz
(L, d)	$(0.94\lambda, 0.75\lambda)$	$(1.15\lambda, 0.92\lambda)$		$(1.41\lambda, 1.13\lambda)$
Array Period	2.0λ	2.5λ	3.2λ	3.0λ
Co-pol Directivity	17.3 dB	19.1 dB	19.5 ± 0.2 dB	20.6 dB
Main-beam location	50°	40°		34°
10-dB beamwidth:				
E-plane	47°	40°	40°	26°
Quasi-H plane	55°	48°	41°	37°
Co-pol				
Main beam efficiency	92%	83%	83 ± 2%	62%

The measured and calculated values for the integrated corner-cube antenna. Measurements at 222 GHz were done with two different ground-plane extensions yielding array periods of 2.6λ and 3.2λ.

elements in a linear array and on single elements surrounded by a ground-plane showed identical results. This can be explained by the array period being larger than 2.5λ and by the vanishing electric-field along the array in a quasi-H plane scan. All patterns shown below are on elements in a linear array environment.

The patterns are competitive with the standard corner-cube antenna with a narrow main-beam and no off-axis sidelobes (Fig. 6). Table I summarizes the salient features of the integrated corner-reflector antenna. The H-plane pattern is wider than expected due to the short extension of the ground-planes from the antenna. The cross-polarization level was lower than -17 dB in the E-plane and quasi-H plane at 222 GHz. The co-polarized main-beam efficiency is calculated from the measured two-dimensional patterns for a reflector illumination of -20 dB. It is defined as the power available in the main-beam divided by the co-polarized power radiated by the antenna. The results show a main-beam efficiency of 92 and 83% for 180 and 222 GHz, respectively. The cross-polarization loss is not expected to be a problem due to the -17 dB measured components in the E- and quasi-H planes. The experimental main-beam efficiencies agree well with the theoretical predictions of Kelly *et al.* [4], Zmuidzinas *et al.* [5], and Grossman [7].

In order to investigate the effect of the ground-plane size on the H-plane pattern, we have fabricated three identical 222-GHz antennas with increasingly larger ground planes. The extension of the ground-planes from the traveling-wave antennas was 0.94λ, 1.6λ, and 2.3λ, yielding an array period of 2.5λ, 3.2λ, and 4.0λ, respectively. The E-plane pattern is not affected by the ground-plane size, but the quasi-H plane pattern narrows considerably between the first two cases and levels off at 41° for an aperture opening greater than 3.2λ (Fig. 7). The resulting pattern for a array period of 3.2λ and $(L, d) = (1.15\lambda, 0.92\lambda)$ at 222 GHz shows a circularly symmetrical mainlobe with E-, quasi-H, and 45°-plane 10-dB-beamwidths of $\sim 40°$ and a directivity of 19–20 dB (Fig. 8). The optimized design is expected to couple well to $f/1.5$-$f/1.8$ imaging systems with an optical coupling efficiency of 75–80% [4], [7].

V. Conclusion

An integrated corner-reflector antenna has been developed for millimeter-wave and terahertz frequencies. A new traveling-wave antenna design, the modified-bend antenna, yields a relatively wide-band input impedance, high-gain patterns between 180–270 GHz, and a -17 dB peak cross-polarization component in the E-plane and quasi-H plane patterns. The monolithic approach allows the integration of a matching

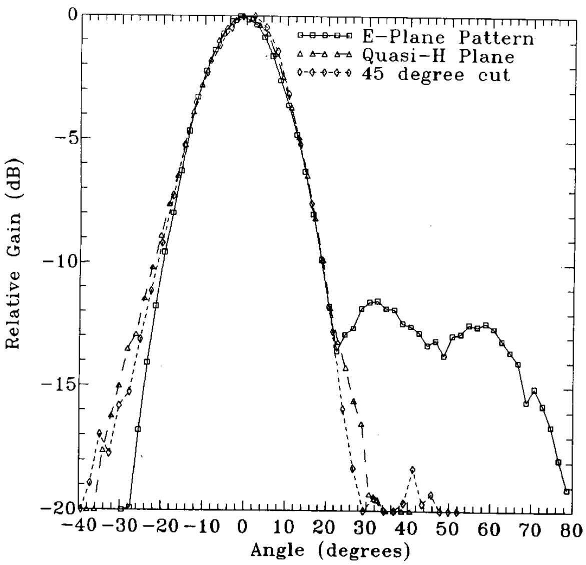

Fig. 8. The measured 222-GHz pattern for $(L, d) = (1.15\lambda, 0.92\lambda)$ and an array period of 3.2λ. The pattern is circularly symmetric with a 10-dB beamwidth of $\sim 40°$.

network and a Schottky-diode or SIS detector at the base of the antenna to yield a low-noise monolithic millimeter-wave receiver, and a 600-GHz integrated corner-cube antenna with a Schottky-diode mixer is currently under development at the University of Michigan.

References

[1] H. R. Fetterman *et al.*, "Far-ir heterodyne radiometric measurements with quasi-optical Schottky diode mixers," *Appl. Phys. Lett.*,, vol. 33, pp. 151-154, 1978.

[2] H. P. Roser, E. J. Durwen, R. Wattenbach, and G. V. Schultz, "Investigation of a heterodyne receiver with open structure mixer at 324 and 693 GHz," *Int. J. Infrared and Millimeter Waves,* vol. 5, no. 3, pp. 301-314, 1984.

[3] E. Sauter, G. V. Schultz, and R. Wohleben, "Antenna pattern of an open structure mixer at a submillimiter wavelength and of its scaled model," *Int. J. Infrared and Millimeter Waves,* vol. 5, no. 4, pp. 451-463, 1984.

[4] W. M. Kelly, M. J. Gans, and J. G. Eivers, "Modelling the response of quasi-optical corner cube mixers," *Instrumentation of Submillimeter Spectroscopy*, E. Kollberg, Ed., *Proc. SPIE* vol. 598, pp. 72-78, 1986.

[5] J. Zmuidzinas, A. L. Betz, and R. T. Boreiko, "A corner-reflector mixer for far- infrared wavelengths," *Infrared Phys.*, vol. 29, pp. 119-131, 1989.

[6] B. Vowinkel, "The main beam efficiency of corner cube reflectors," *Int. J. Infrared and Millimeter Waves,* vol. 7, p. 155, 1986.

[7] E. N. Grossman, "The coupling of submillimeter-wave corner-cube antennas to Gaussian beams," *Infrared Phys.*, vol. 29, p. 875, 1989.

[8] M. Dye, "Design of a 75-110 GHz MMIC mixer incorporating the surface channel Schottky-diode," M.S. thesis, Univ. Virginia, Charlottesville, May 1990.

[9] H. P. Roser, R. Wattenbach, E. J. Durwen, and G. V. Schultz, "A high resolution spectrometer for 100 μm to 1000 μm and detection of CO (J = 7.6), CO (J = 6-5) and CO (J = 3-2)," *Astron. Astrophys. J.*, vol. 165, 287-299, 1986.

[10] R. T. Boreiko and A. L. Betz, "Heterodyne spectroscopy of the J = 17-16 CO line in Orion," *Astron. Astrophys. J.,* vol. 337, pp. 332-341, Feb. 1989.

[11] S. S. Gearhart, C. C. Ling, and G. M. Rebeiz, "Monolithic millimeter-wave corner-reflector antennas," presented at *IEEE Symp.*, Dallas, TX, May 1990.

[12] S. S. Gearhart, C. C. Ling, and G. M. Rebeiz, "Integrated 222 GHz corner-reflector antennas," *Microwave and Opt. Tech. Lett.*, Special Issue on Space Terahertz Technology, Jan. 1991.

[13] K. E. Peterson, "Silicon as a mechanical material," *Proc. IEEE*, vol. 70, pp. 420-457, 1982.

[14] R. E. Williams, *Gallium Arsenide Processing Techniques*. Dedham, MA: Artech House, 1984.

[15] R. S. Elliott, *Antenna Theory and Design*. Englewood Cliffs, NJ: Prentice-Hall, 1981.

[16] S. S. Gearhart, C. C. Ling, G. M. Rebeiz, H. Davée, and G. Chin, "Integrated 119-μm linear corner-cube array," *IEEE Microwave and Guided Wave Lett.*, vol. 1, pp. 155-157, July 1991.

[17] D. P. Neikirk, W. W. Lam and D. B. Rutledge, "Far-infrared microbolometer detectors," *Int. J. Infrared and Millimeter Waves,* vol. 5, no. 3, pp. 245-277, 1984.

Steven S. Gearhart (S'91) was born in Roanoke, VA, in July 1965. He received the B.S.E.E. degree from Virginia Polytechnic Institute and State University, Blacksburg, in 1988 and the M.S. degree in electrical engineering from the University of Michigan, Ann Arbor, in 1990, where he is currently working toward the Ph.D. degree in electrical engineering.

His research interest include millimeter-wave antennas, monolithic receivers, and solid-state fabrication.

Curtis C. Ling (S'90) received the B.S. degree in electrical engineering from the California Institute of Technology, Pasadena, in 1988, and the M.S. degree in electrical engineering from the University of Michigan, Ann Arbor, in 1990, where he is currently enrolled in the Ph.D. program.

His research is in terahertz and millimeter wave integrated antennas and circuits. He is working on a 94 GHz monolithic monopulse radar design, planar subharmonic receivers as well as wide-band THz power measurement.

Gabriel M. Rebeiz (S'86–M'88) was born in Beirut, Lebanon, in December 1964. He received the B.S. (honors) degree from the American University in Beirut, Beirut, Lebanon, in 1982, and the Ph.D. degree from the California Institute of Technology, Pasadena, in 1988, both in electrical engineering.

He joined the faculty of the University of Michigan, Ann Arbor, in September 1988 where he is now an Assistant Professor in the Electrical Engineering and Computer Science Department. His research interests lie in planar millimeter-wave antenna structures, monolithic receivers, and fabrication and measurement of novel millimeter-wave transmission-lines and devices.

Dr. Rebeiz has invented a high-efficiency integrated-horn antenna (U.S. Patent) and has been awarded a NASA-Certificate of Recognition Award for his contribution to the millimeterwave space program (March 1990) and the best paper award at the 1990 International Conference on Antennas, Nice, France. He received an NSF Presidential Young Investigator Award in 1991.